CLUTTER SUPPRESSION IN RADARS BY AN APPLICATION OF DIGITAL FILTERS TO MTI SYSTEMS

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THESIS

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BY AN APPLICATION OF
DIGITAL FILTERS TO MTI SYSTEMS

by

John Elias Sarris

December 1979

Thesis Advisor: J. Bouldry

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This thesis presents the significant aspects of old, improved and modern anticlutter techniques, each of which is described with its principal advantages and disadvantages.

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Continuing, an application of digital filters is made by designing a fourth-order recursive filter with Butterworth characteristics, using a computer program. This filter can be used in a range-gated MTI (Moving Target Indication) receiver to improve its performance.

Special consideration was given to the AN/UPS-1 and AN/SPS-40B,C radars which are used in the Greek Navy in a clutter environment characteristic of the Mediterranean Sea. Appropriate modifications to their existing analog systems are suggested.

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Clutter Suppression in Radars By an Application of Digital Filters to MTI Systems

by

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I. INTRODUCTION

The conglomeration of unwanted echoes coming from fixed targets is defined as "clutter". These echoes clutter the radar display and make the recognition of wanted signals difficult. The result degrades the quality of radar information.

There are several types of clutter such as those caused by weather, ground, sea, birds, chaffs, rain, clouds, precipitation and other meteorological phenomena. In addition, second-time-around clutter is often troublesome.

For these types of clutter, there are a number of remedies for improving the signal-to-clutter ratio. Consequently radar systems are configured using one or more of these methods.

The basic signal parameter which is used for distinguishing fixed and moving targets is the doppler shift in frequency of the moving target. This frequency shift is given by the relationship

$$f_d = \frac{2V_r}{\lambda} \tag{1.1}$$

where V_r is the radial relative velocity component of the target and λ is the wavelength of the transmitted signal.

The presence of various frequencies is the basis for the use of a filter that passes echoes with a doppler shift but rejects clutter which has zero or small doppler shift.

Over the last twenty years, a concerted effort has been made to understand the root causes of clutter, its characteristics, and to find ways for MTI performance to be improved. This effort has resulted in new signal processing techniques, which together with certain antenna arrangements, minimize clutter problems.

Analog MTI is one technique which has been used in radar systems for a number of years by the Greek Navy with limited success. Performance has suffered from the inherent difficulty of controlling the delay time and system gains. The on-line availability has been generally poor. The radars operate well when tuned-up, but experience deteriorating performance with time. The effective contribution to the overall radar system performance has generally been limited due to the inflexibility of the hardware.

However, many of the disadvantages of the analog MTI have been eliminated by digital MTI. This is shown in this thesis using computer programs to design an optimum filter. No attempt was made to implement the filter with hardware. However, the computer results are in a form consistent with the format needed for easy digital filter design.

It is expected that when the entire system is built and operated, it will have the inherent flexibility and stability required to obtain greatly improved performance against clutter.

II. OLD TECHNIQUES

In the design of the original radar systems, the main effort was directed toward the ability of the radar receiver to detect the presence of echoes and extract information from them.

Fundamentally, detection was limited by the presence of noise in the receiver. In addition detection of moving targets was inhibited by the presence of clutter frequently much stronger than the moving targets themselves.

Under such circumstances the radar became useless. The need to overcome this led to the design of additional analog subsystems (techniques), which for that time, offered considerable detection improvement. A number of these could be mentioned, but those few which are still employed in older but existing radar systems will be described.

Part II covers the following techniques.

(i) Frequency Agility
(ii) Circular Polcarization
(iii) Sensitivity Time Control - Logarithmic Receivers
(iv) Instantaneous Automatic Gain Control

A brief description of each is given, indicating the advantages and disadvantages.

A. FREQUENCY AGILITY

Frequency agility is a technique which enables one to achieve a measure of clutter reduction by changing the

frequency. Coaxial magnetrons permit a change in the basic frequency by mechanical means. Signal enhancement in clutter by a frequency change is shown in Figure 1 which is an A-scope presentation of a rooftop radar [Refs. 1,2].

Part (a) of Figure 1 [Ref. 1] is with the agility switch ON and part (b) OFF. The radar was operated to look at small targets over a fairly cluttered ground terrain. In this case the bandwidth of agility was about 75 MHz. Clearly a distinct improvement is seen and the two hard-targets are more easily discerned.

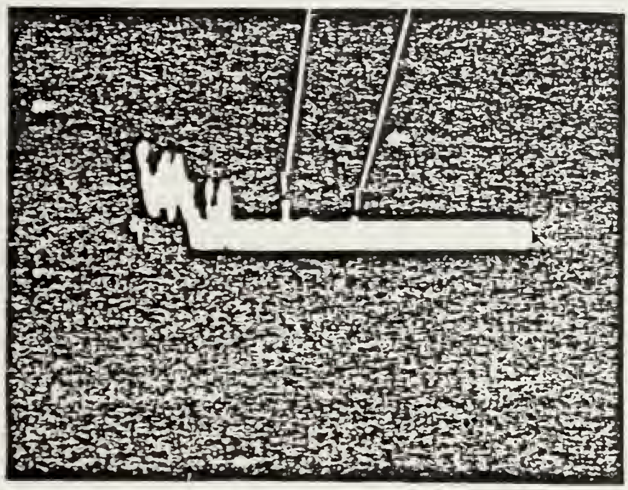
If there is a doubt about the presence of targets, then a quick flip of the Agility switch from ON to OFF can usually help resolve the doubt.

To obtain frequency agility mechanical adjustments are made within the microwave tube. These adjustments may be as follows:

(i) By moving the end wall of the coaxial cavity up and down slowly. This movement results in broad-band tuning.

(ii) By varying the ring's diameter of the outer periphery of the cavity. This form is independent of (i) and causes changes in the tuning sensitivity of the cavity.

The second technique gives a tuning speed ranging from 80 to 200 Hz/second and a tuning bandwith up to 100 MHz.



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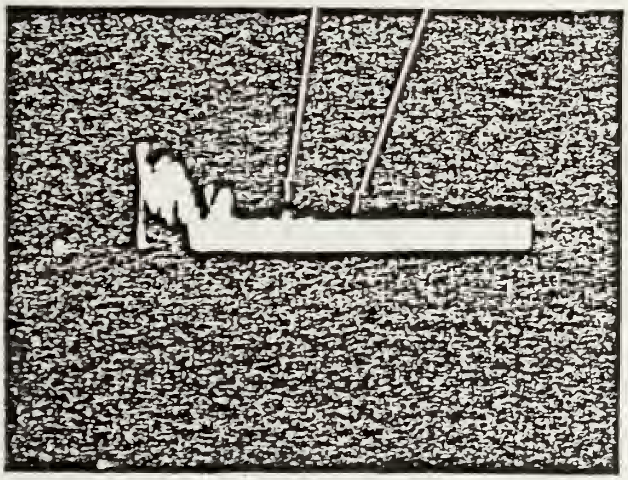


1(b)

Figure 1. A-scope presentations, (a) with and (b) without frequency agility.

The advantage of this type of tuning is the improved performance as demonstrated by its use in operational systems today.

However, there is a disadvantage, when precipitation extends over a large volume and completely fills the antenna beam. Then, the total energy received from all the precipitation particles within the radar resolution cell can be quite large compared with that received from a conventional target and can mask or clutter the radar display.

In this case the amount of clutter with which the target echo must compete may be reduced with narrow beam-widths and narrower pulse widths if otherwise acceptable.

The best method, which can be applied to any radar system, of eliminating weather effects, such as heavy precipitation, is to operate at lower radar frequencies. It may be noted that in weather radar applications, the designer has the option of operating at the lower frequencies which give a better S/C ratio.

B. CIRCULAR POLARIZATION

When a radar operates at the higher frequencies and the target is of a complex nature like an aircraft, it is possible to improve the detection capability in the presence of precipitation by radiating circularly polarized energy [Ref. 3].

Successful detection is based on the dissimilarity between clutter and aircraft echoes. With precipitation

and raindrops, the backscattered energy does not depend on the polarization of the incident waves. But in the case of the aircraft target the energy is dependent on the incident polarization. For example, the radar may use an antenna which is responsive to rotation of the polarized wave in one direction. Since symmetrical reflectors reverse the rotation of the wave, that echo energy will not be accepted by the receiver.

Circular polarization is generated by two methods as outlined in Ref. 3 and as shown in Figure 2(a,b). The first part shows a quarter-wave plate which converts linear polarization to circular by passing the wave through it. The second part shows a turnstile junction which is a combination of rectangular and circular waveguides properly matched to generate circular polarized waves.

Generally with circular polarization the ratio of target to precipitation signal is improved by 8 to 25 dB over that obtained with linear polarization. Under normal conditions an improvement of 15 to 20 dB is expected. It must be noted that in all previous cases the ground reflections are considered negligible.

C. SENSITIVITY TIME CONTROL - LOGARITHMIC RECEIVER CHARACTERISTICS

1. Sensitivity Time Control (STC)

One of the earliest and most widespread but less sophisticated techniques for attenuating the clutter signal

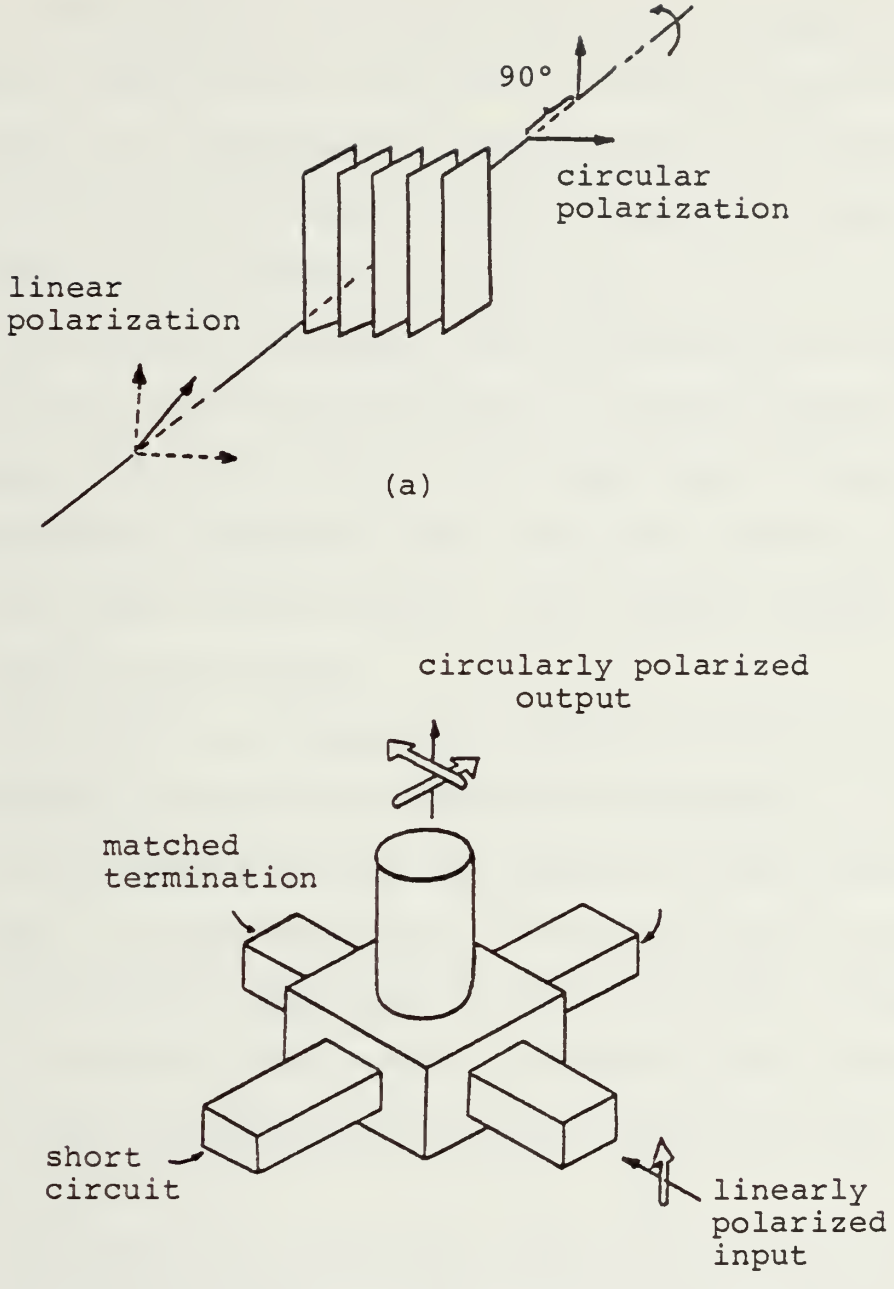


Figure 2. Quarter-wave plate and turnstile junction for converting linear to circular polarization

is the gain at close ranges. This is known as STC [Refs. 2, 3, 13]. The reduction of clutter from close targets is achieved by reducing the likelihood of receiver saturation, which leads to a programmable receiver gain as a function of time so that the gain will be low for targets at close range and high for targets at long range.

The problem of dynamic range limitation of radar receivers was overcome by switching on the STC circuit. This simply reduced the gain of the amplifiers, usually at the intermediate frequency stage. The gain varies as R^n, where n can take a value between 2 for weather effects and 4 for land and sea clutter.

The resulting improvement appears on the PPI, where the STC circuit reduces the excessive brightness at the center of the display due to the bunching of the radial sweep lines. This anti-clutter circuit is used only where needed and should be switched off at other times. Consequently, the use of this technique depends upon an operator's judgment and is more helpful the more experienced the operator.

In addition to the above, this type of anti-clutter technique has the limitations caused by its analog circuitry and IF stage processing, which limit the radar's clutter reduction and target detection capabilities.

Similar to this is the fast-time constant (FTC), another analog circuit. This is normally a high-pass

filter (RC), as outlined in Ref. 3. Its use also depends on the judgement of the radar operator. It is commonly used in conjunction with STC and is effective on all targets including clutter at all ranges of the radar.

2. Logarithmic Receiver Characteristics

The receiver gain, when clutter echoes are present, can be reduced to prevent relatively large clutter signals from saturating the receiver by using a logarithmic amplifier followed by a differentiating circuit or a pulse-length discriminator [Refs. 2, 3].

This technique is used to reduce the effective dynamic range of clutter or jamming signals at the threshold detector or display. Since clutter varies in intensity with range, the use of a log-receiver reduces this and provides a clutter-signal amplitude independent of range at the output.

The transfer characteristic of a log-amplifier has the form

$$y = a \ln(1 + bx) \tag{2.1}$$

where a, b are constants of the amplifier, x is the instantaneous amplitude of the input and y is the output [Ref. 2].

The purpose of the differentiator circuit, which follows the log-amplifier, is to reduce the mean value of the noise return or clutter. This is successful only when

the clutter has a uniform distribution in range for a period greater than that of the FTC circuit.

At this point, techniques 1 and 2 are compared. Sensitivity Time Control (STC) suppresses clutter echoes, which enter via the sidelobes much better. STC automatically turns down the gain at close ranges, reducing in that way sidelobe signals. The log-amplifier does not suppress side-lobe clutter which masks the radar display. Both techniques have the disadvantage of requiring perfect adjustment and skilled operators.

There are a few applications where these two techniques may be employed together. Marine radars use log-receivers exclusively because they permit strong and weak echoes (returns of sea, buoys and boats) to appear on the radar scope.

The block diagram of Figure 3 corresponds to a log-amp followed by a pulse-length discriminator (PLD). This form of receiver is an alternative solution having the same effect on clutter and noise as the log-FTC circuitry, rejecting, as well, forms of land clutter and pulsed interference.

The bandpass filter plays the role of the matched filter of the conventional linear receiver, and passes pulses within fixed frequency limits which are equal to the pulse period of the delay line.

The disadvantage of this technique is the same as the log-FTC, in that loss of detectability results. These

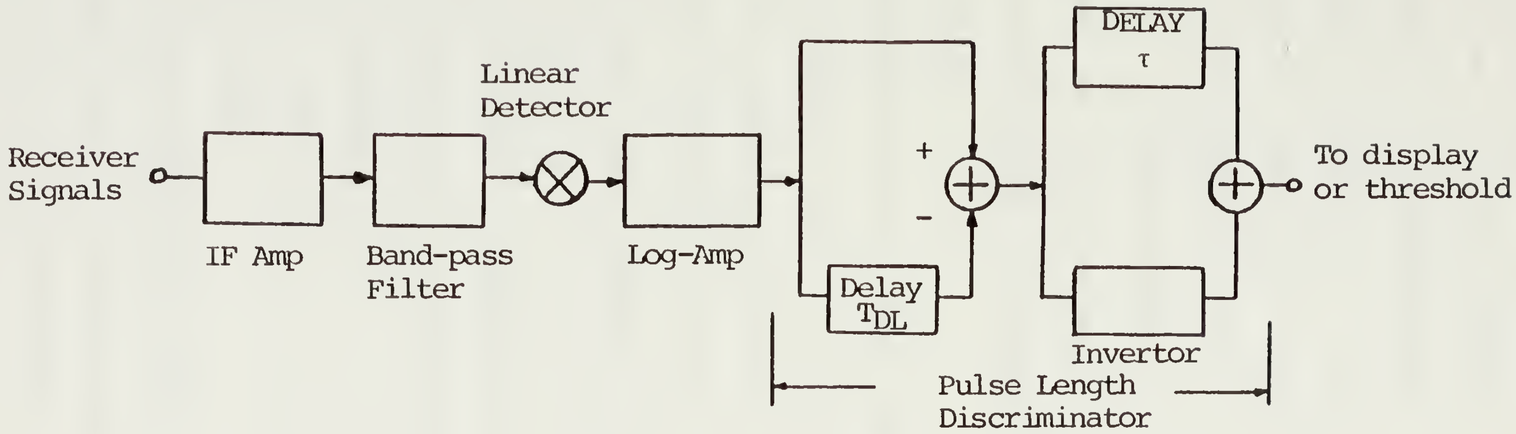


Figure 3. Log-amp pulse-length discriminator receiver

losses reach the level of 4-8 dB for single-hit detection and 2-4 dB for u-Pulse postdetection integrations.

D. INSTANTANEOUS AUTOMATIC GAIN CONTROL (IAGC)

The IAGC technique is a fast-acting form of AGC to prevent the saturation of IF amplifiers. It is used in most radar receivers and acts in a similar manner to the automatic volume control (AVC) in a radio receiver.

A block diagram is shown in Figure 4, with greater detail given in Ref. 3. The control of the gain of the IF amplifier is obtained by using negative feedback. Its filter has a narrow bandwidth to minimize the effect of the extended clutter echoes.

There is a limitation on the IAGC loop, which can only control one or two stages. For this reason instabilities sometimes appear in a system with more stages.

A second technique similar to IAGC is that known as detector balanced bias (DDB), which is shown in the diagram of Figure 5. DDB differs from IAGC in three basic respects:

(i) Does not employ a feedback loop.

(ii) Bias is applied to a diode detector instead of the IF amplifier.

(iii) Employs a delay line in the absence of the feedback loop.

While some of these techniques used singularly or in combination are beneficial in reducing the effects of some types of clutter, they are not sufficient in modern

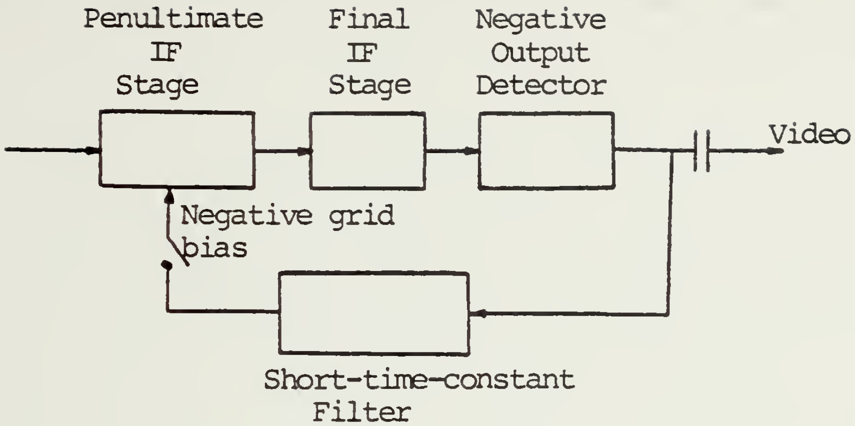


Figure 4. Instantaneous Automatic Gain Control (IAGC)

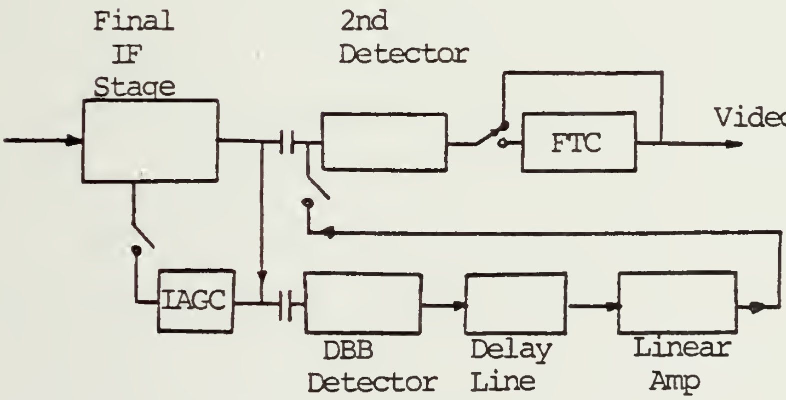


Figure 5. Detector balanced bias (DDB)

sophisticated radar systems. More importantly they are not effective in reducing the types of clutter that occur in the Mediterranean environment which is the concern of this thesis.

III. IMPROVED TECHNIQUES

In World War II a particular class of radar processors made its appearance. This type of processor extracted from the echo the doppler shift in frequency caused by a moving target. Originally, such radar systems employed CW transmission. The newer systems were able to discriminate between the echoes of moving and fixed targets.

At that time the problem of clutter suppression was approached from many directions; "clutter" meaning the return echoes from fixed targets in the vicinity of the target of interest. Sea, ground, weather effects, angels, birds, rain, chaff, trees and vegetation movement are common sources of clutter.

The analog techniques originally developed usually employed delay line cancelers or analog filter banks (matched). These techniques are still used in large numbers in contemporary applications. Delay line cancelers are used in the UPS-1 and AN/SPS-40B,C radars on ships of the Greek Navy.

Thus, this thesis continues to compare the performance of other techniques, both existing and modern, in an effort to find a satisfactory clutter rejection filter that is not unduly complex and can be rather easily implemented.

Part III describes the following representative techniques with their respective advantages and disadvantages.

(i) Analog MTI and Pulse-Doppler Systems.

(ii) Dual Beam Antenna.

(iii) Log - CFAR.

(iv) Video Detector.

A. ANALOG MIT AND PULSE-DOPPLER SYSTEMS

When doppler information is extracted from a pulsed radar return, it is known as an MTI radar. The MTI stands for "moving-target indication". It may also be called pulse-doppler radar. In practice, a distinction is sometimes made between the MTI radar and the pulse-doppler radar, although they are both based on the same principle.

MTI usually refers to a radar in which the doppler frequency measurement is ambiguous but the range measurement is unambiguous. Another characteristic feature of the MTI radar is the delay-line canceler used to detect the doppler frequency shift.

In the pulse-doppler radar the doppler measurement is usually unambiguous and the range may or may not be ambiguous. Ambiguous range means that multiple-time-around echoes are possible, while ambiguous doppler implies that "blind speeds" fall within the range of expected target speeds.

MTI radar has the ability of extracting the moving-target echo from the clutter echo even when the clutter is 20 to 30 dB greater than the moving-target echo. Some pulse doppler radars can detect moving targets even when the clutter echo is 70 to 90 dB greater than the target echo.

The distinction between the two radars arose historically, and is commonly still made. In many instances the difference between them is only a matter of nomenclature.

Figure 6 shows the simplest MTI processor, the single delay-line canceler, which extracts the doppler information by subtracting two successive echoes. The delay-line canceler plays the role of a filter, eliminating the d-c component of fixed targets and permitting the a-c components of moving targets to pass.

There are several types of MTI processors. A distinction between them is the type of information processed in the returned signals, that is, whether the phase, the amplitude, or both phase and amplitude are processed. These types are described with their block diagrams in Refs. 2 and 3. It is beyond the stated intention of this thesis to describe all types of MTI, but some are listed here for reasons of interest.

(i) Phase-Processing MTI

(ii) Amplitude-Processing MTI

(iii) Vector-Processing

(iv) Clutter-Locking MTI

(v) Multiple-Canceler Systems

These processors do not satisfactorily handle the well known problems of clutter precipitation. This problem becomes more important when the effects of wind shear and average clutter velocity are taken into account. For example,

the clutter performance of the aforementioned three radars is poor under these conditions. This has been verified by the experience of the Greek Navy in the Mediterranean environment.

To get a clearer idea of this limitation on MTI performance, the clutter improvement factor I was plotted vs. range. Figure 7 shows the plot. The equation of I is as follows

$$I \doteqdot \frac{\lambda^2 f_r^{\,2}}{8^{\,2}(\sigma_{turb}^2 + .18k^2R^2\phi^2)} \qquad (3.1)$$

and the Continuous System Modeling Program 1 was used with some typical values given:

σ_{turb} = variance of wind turbulence, varying from .7 to 1.1 meter/sec

λ = .30 m, transmitted wavelength

f_r = 600 Hz, PRF

k = 5.7 meter/(sec)(km), is the component of wind velocity gradient in the direction of the radar beam

R = range in km

ϕ = 1.5°, vertical half-power beamwidth-two-way path in radians.

Figure 7 shows that the attenuation of clutter is quite good for ranges up to 18 km where, for σ_{turb} = .7 and

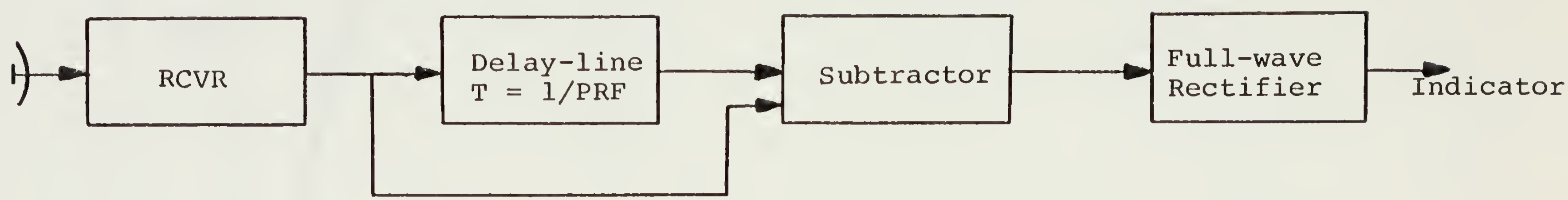


Figure 6. MTI using delay-line canceler

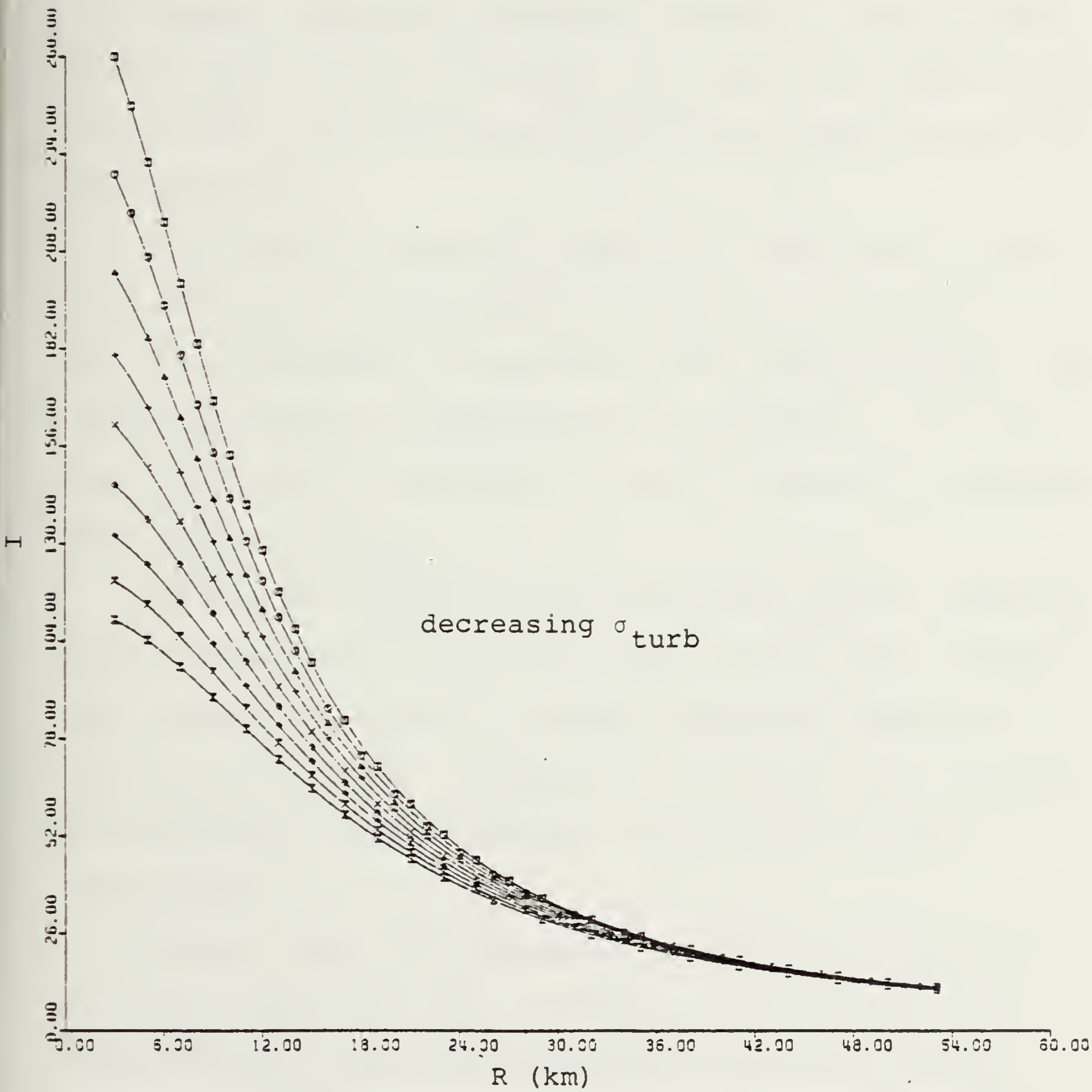


Figure 7. Clutter improvement factor vs. Range

1.1 meter/sec a difference of 1.5 dB is measured. As range increases, the clutter attenuation decreases rapidly as indicated.

B. DUAL BEAM ANTENNA

The MTI visibility is highly degraded at short ranges due to nearby graound clutter, sea clutter and angel echoes. To eliminate these problems a dual beam antenna system has been designed.

This system is shown in Figure 8. Using the low and high beam horn at the appropriate time, as outlined in Ref. 13, the clutter is reduced at the receiver input. The change in the use of horns occurs by switching. The two horn noses have a difference of about 4 degrees in elevation. The angle is variable.

The degree of clutter suppression and target visibility can be determined by the beam angle and switching distance. The coverage is varied by varying the angle difference in the nose directions. Consequently, the switching distance is changed and this determines the degree of clutter suppression.

Figure 9 [Ref. 13] compares the S/C ratio of dual beam antennae with single beam antennae. After determining target/clutter ratios by a computer simulation program, conclusions from these two techniques are obtained. This comparison shows that a better improvement factor is obtained using a dual beam antenna system.

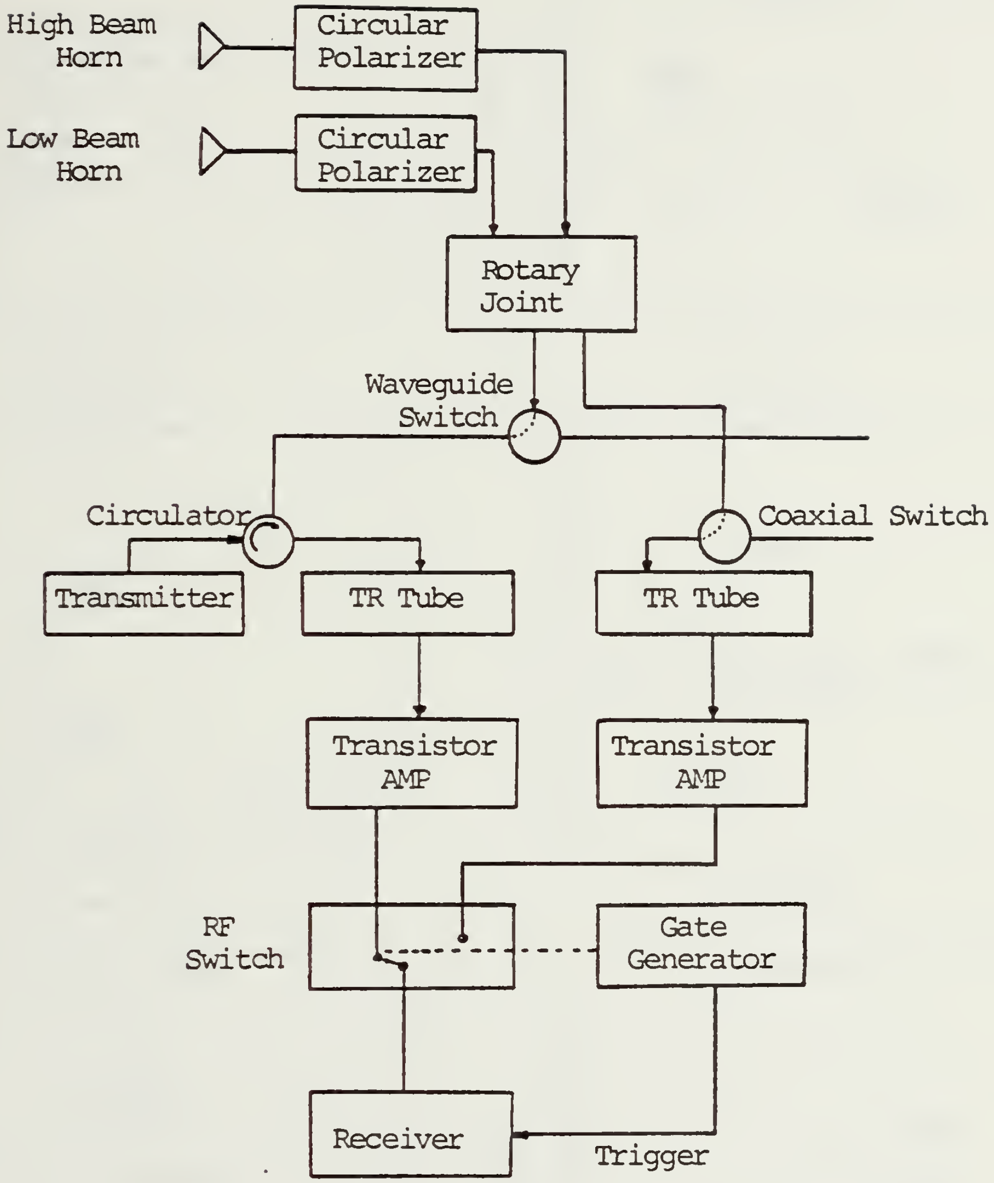
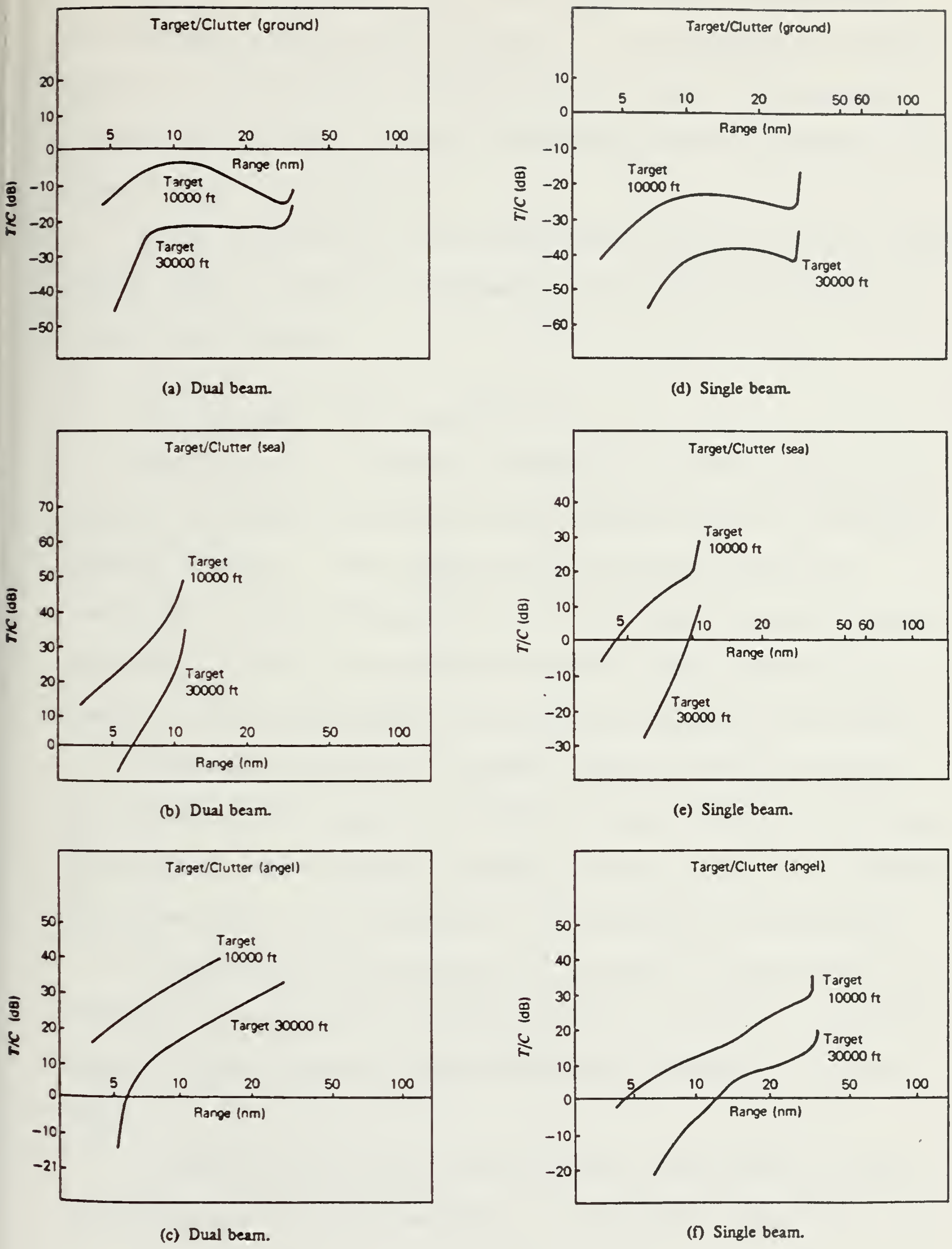


Figure 8. Dual beam antenna system block diagram



(a) Dual beam.

(b) Dual beam.

(c) Dual beam.

(d) Single beam.

(e) Single beam.

(f) Single beam.

Figure 9. Comparison in S/C ratios between dual beam and single beam antenna systems.

Types of clutter, which are taken into account here are ground, sea and angel echoes. In all cases an improvement of 15 to 20 dB is obtained using dual beam antennas. Detection, however, becomes difficult when the range is less than 5 nautical miles.

These advantages of the dual beam antenna are the reasons which make it useful in systems such as ARSR (Air Route Surveillance Radar).

C. LOG-CONSTANT FALSE ALARM RATE

Additionally, a special technique for ARSR systems is that of Log-CFAR, which has been widely used to eliminate weather clutter. The reason this technique has been employed is that neither MTI nor dual beam antenna systems have the ability to suppress weather clutter which has a radial velocity distribution at high angles of elevation.

Previously mentioned methods obtain some improvement by improving the signal-to-clutter ratio. But in the case of clutter, whose signal strength is well above the threshold or noise level, the output of the receiver is saturated.

An additional suppression of clutter is needed and is obtained by using the Log-CFAR method. It keeps a constant output at the receiver suppressing the clutter to noise level.

In terms of the block diagram this technique is shown in Figure 10 and its operation is outlined in Ref. 13. A

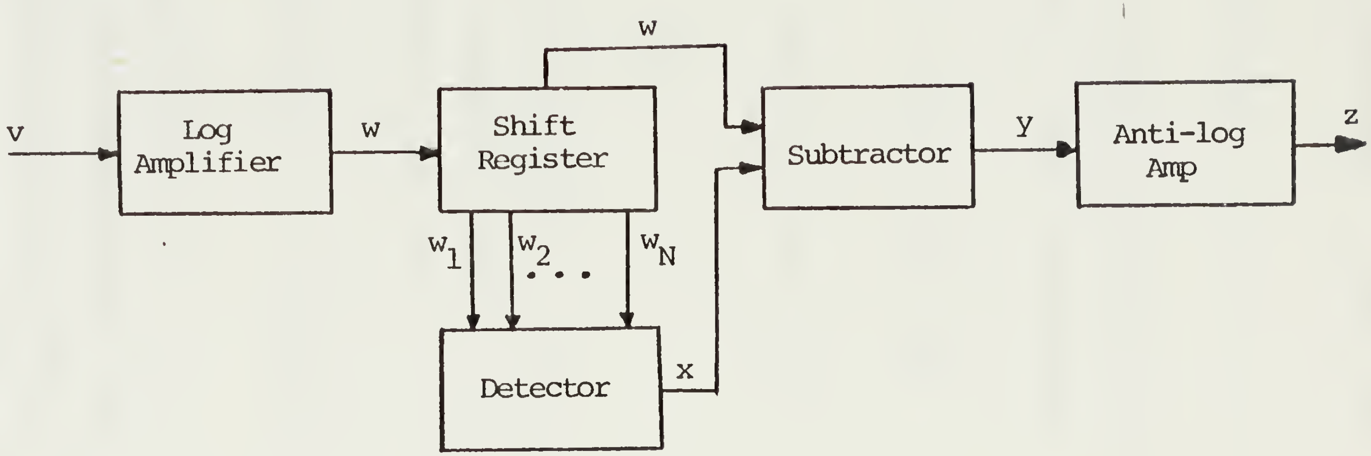


Figure 10. Log-CFAR system block diagram

step by step explanation is given in terms of input-output waveforms according to the diagram of Figure 10.

The output w of log-amplifier is given by

$$w = a \ln(1 + bv) \qquad (3.8)$$

where a,b are constants and v is the input [II-C-2]. The subtractor has an output y, which is given by

$$y = w - x = a \ln \left(\frac{v}{\sigma}\right) + \frac{a\gamma}{2} \qquad (3.9)$$

where x is the average detector value given by

$$x = \frac{1}{N} \sum w_i \qquad (3.10)$$

and where γ is the Euler's constant, σ the variance of the input signal v and N, the sample number. The output signal of the anti-log converter is given by

$$z = c \exp (d \cdot y) = \frac{cv}{\sigma} \exp \left(\frac{\gamma}{2}\right) \qquad (3.11)$$

where c,d are constants and $a \cdot d = 1$.

Sample number, N, plays an important role, since it determines the processing loss of the CFAR circuit and consequently the clutter suppression. Reference 13 shows that the cutoff frequency of the rejection band is given by

$$f_c = \frac{1}{N\tau} \tag{3.12}$$

where τ is the pulse width.

The Log-CFAR process has two disadvantages which are listed below.

(i) Sample number N must have a value of about 8 to keep the processing loss between 1 and 2 dB.

(ii) Mean value of clutter component must have a frequency less than $1/N\tau$ to be eliminated.

D. VIDEO DETECTOR

This application is specially useful in non-coherent airborne radars operating either over land or in the proximity of a land mass, where very intense ground clutter patches or individual scattering objects exist.

Such ground clutter may have radar cross-sections on the order of one to ten million square meters. Thus, even with the best designed antennas on airborne platforms, this clutter may be strong enough to be detectable through the antenna sidelobes.

These signals in general will have doppler velocities within the acceptance band of the MTI filter or filter bank and will result in a high rate of false alarms. The side-lobe clutter problem is a very serious one but the use of a square-law detector provides significantly improved clutter rejection.

For main lobe clutter which is assumed to be Gaussian in both amplitude and frequency, the MTI improvement factor is calculated for different spectral widths using linear and square law detector characteristics.

The block diagram of the system is shown in Figure 11 where a magnetron is used as the transmitter. The echo signal is converted to IF and then amplified. Detection is realized by the so called Video Detector. After MTI filtering, digital signal processing follows before the video presentation on a display.

The signal x(t) has a general spectrum in the IF shown in Figure 12. Assuming that the clutter component is much larger than the other components, the video spectrum is shown in Figure 13. After the video detector this spectrum is approximately the auto-convolution of the input spectrum. Convolutions not containing C, which stands for clutter, are neglected from the spectrum. Component C*C is undesirable clutter and is suppressed by a digital high-pass filter.

The system described above, using the square law detector provides a wider region of good clutter suppression. It is clearly superior in comparison with other improved clutter suppression techniques, but the dynamic range of the output signal is, of course, larger for the square law detector which is a disadvantage in several applications. It can be noted that at ranges where there is no clutter the target is not detectable after the MTI filter. For target detection, the filter must be bypassed at such ranges.

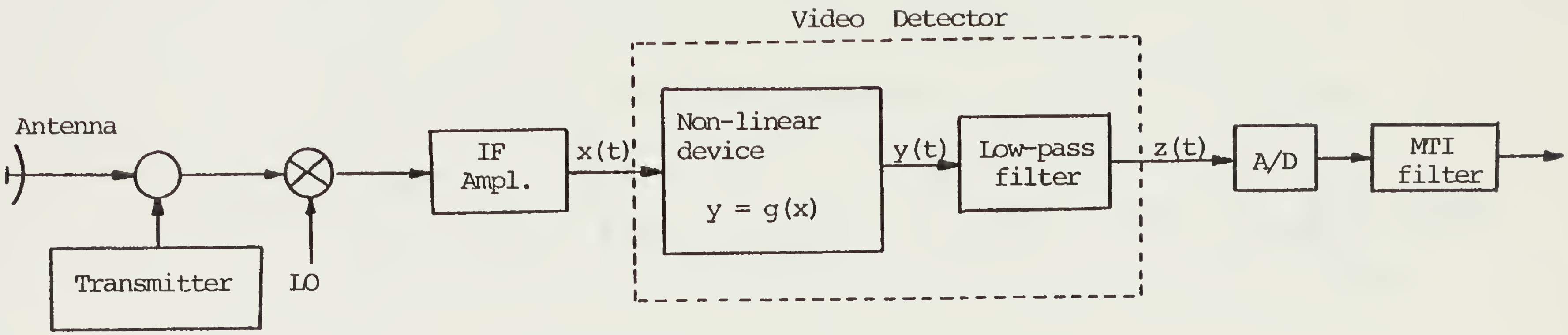


Figure 11. Non-coherent radar system block diagram

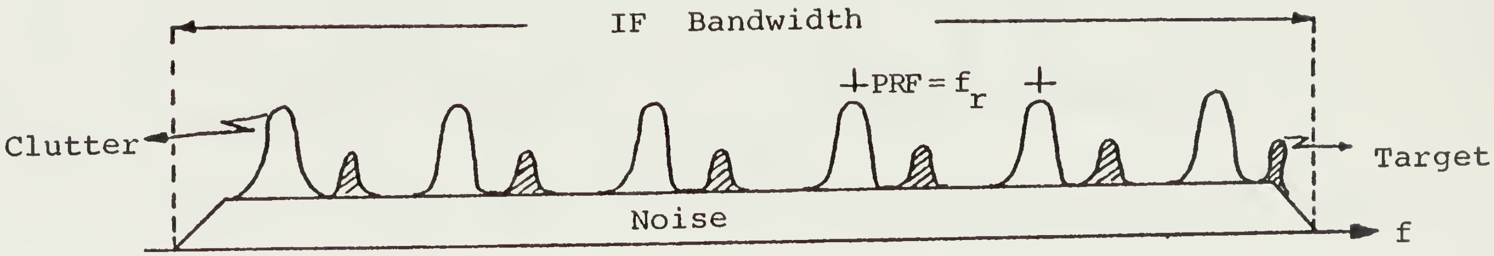
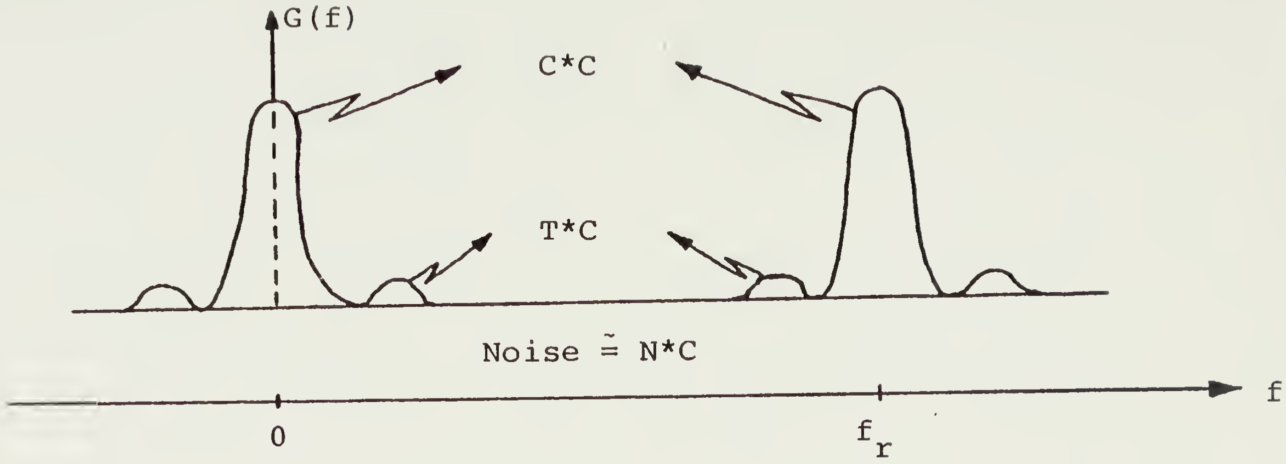


Figure 12. Power spectrum of IF signal x(t)



* = Convolution, T = target, C = clutter, N = noise

Figure 13. Power spectrum of video signal z(t)

Clutter beyond the unambiguous range, after convolution with first time around echoes, results in white noise. This is an additional system problem.

IV. MODERN TECHNIQUES

This section is prepared to demonstrate several "modern techniques" which are employed in radar system technology. The word "modern" covers the developments of the past 5 years and includes future trends in signal waveform design and processing technology. It describes three techniques, which are listed below.

(i) Pulse Doppler Processing

(ii) Moving Target Detector

(iii) Application of FFT Algorithm.

Finally, a tabulation is made to compare the amount of clutter suppression achieved by each technique in the three categories on different types of clutter.

A. PULSE-DOPPLER PROCESSING

Here a technique is presented which is especially applicable to high-power, pulse-doppler radars, achieving good clutter discrimination. The basis of this method is an adaptable filtering and integration system which processes the signal in a way that the processin loss is minimized. Besides, good range resolution and velocity resolution can be obtained. The integration is made adapative to the changes in the variable PRF by using commutated-capacitor filters.

The clutter filter system has a block diagram shown in Figure 14. The signal waveform looks like a Gaussian-modulated envelope due to its antenna characteristics. A Gaussian filter is placed at the beginning, which broadens the duration, τ, of the pulses and consequently reduces their peak amplitude.

This filter provides a narrow spectrum of the signal for adjacent channel rejection during the mixing processes. Then a Gaussian low pass filter rejects the fundamental frequency f_o and all harmonics generated by the previous process. After the sample and hold circuit a high pass filter is connected. The purpose of this filter is to attenuate clutter components to below noise level.

The two signals are then mixed with f_o. Since these are out of phase by $\pi/2$ radians, they are combined to provide an output signal from which clutter has been removed.

Some important advantages of this technique are listed below [Ref. 16].

(i) Clutter suppression at levels of about 80 dB down in amplitude.

(ii) Use of capacitors in association with adaptable sample and hold circuits, stationary clutter is reduced.

(iii) Rapid changes of PRFs during the time on target permits the system to respond without spreading the clutter or destroying the integrity of the signal information.

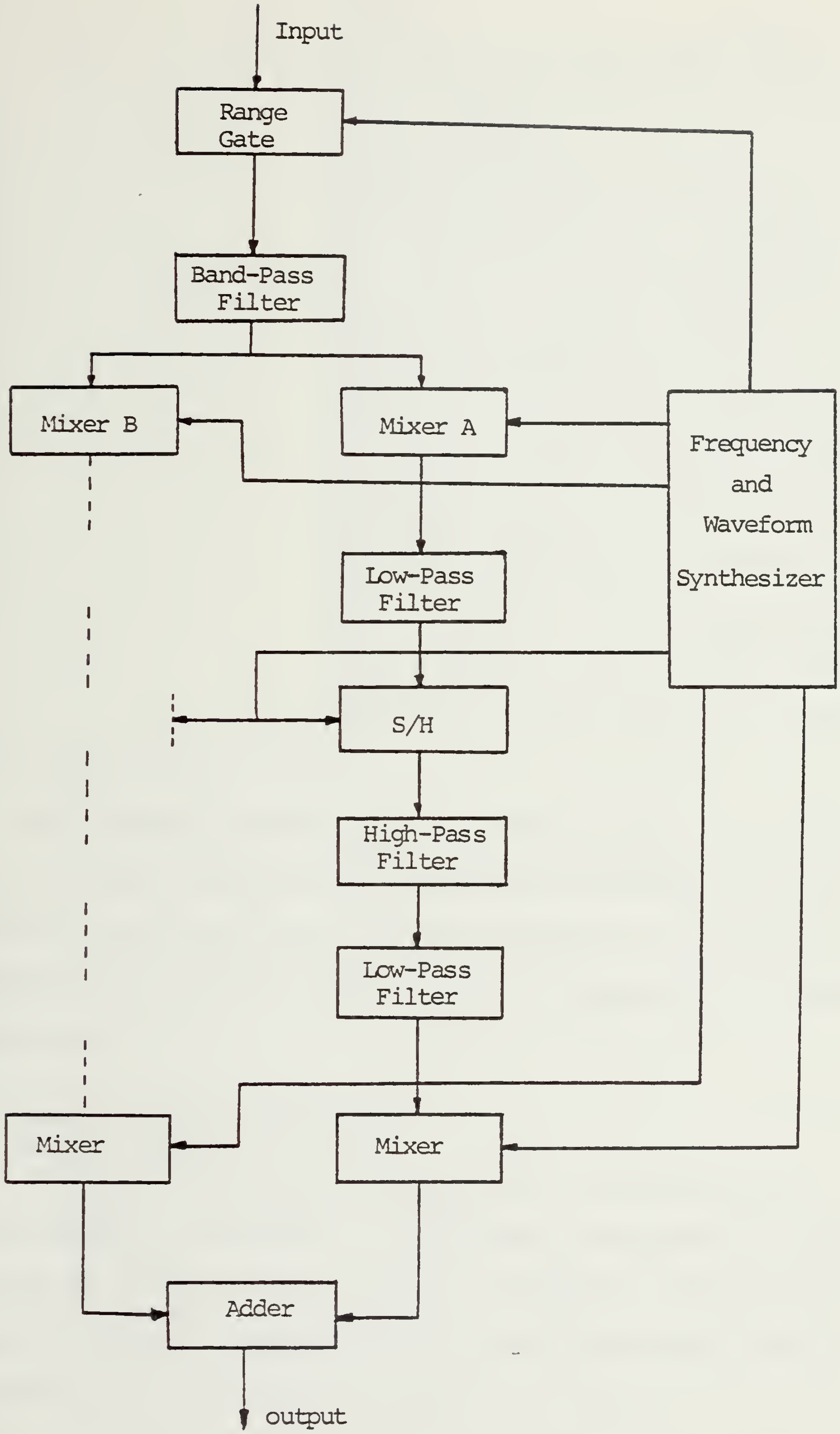


Figure 14. Clutter filtering system

(iv) The low-pass filter reduces system noise to a much lower level.

The disadvantages of the system are the range and velocity ambiguities which occur when it is required to detect very fast targets. Also, the greater complexity due to three distinct stages of spectrum analysis.

This study leads to the conclusion that the introduction of clutter filters, which are switched in synchronism with the pulse repetition frequency sequence, effectively reduce the generation of clutter side-bands by an additional 40-50 dB over a fixed filter system. It is thus possible to design a long range Pulse Doppler Radar System with more than 80 dB clutter suppression, if the PRFs can be switched at a rapid rate.

The storage elements are capacitors which are switched by a hardware program to integrate or discharge signals. By this means real-time target assessment is realized. The performance of this radar system is consequently significantly improved.

B. MOVING TARGET DETECTOR

Recently, the Lincoln Laboratory, associated with Massachusetts Institute of Technology, developed a signal processor named the Fast Digital Processor which performs a Digital Fourier Transform on the received signal and presents the Fourier coefficients [Refs. 1, 17, 18].

This processor is very fast, since it is a special-purpose computer designed to perform signal processing tasks such as FFTs. Its algorithm was tested by embodying it in a hardwired version of the Moving Target Detector (MTD). The MTD has been designed, built and tested at Lincoln Laboratory where it is now undergoing further tests.

The MTD technique uses the cascade of an MTI and pulse doppler processor to achieve clutter rejection. This combination is an approximation to an optimum clutter rejection filter.

In terms of the block diagram, the MTD is shown in Figure 15, where it is seen that the signal to be processed arrives at the IF stage from the output of the IF pre-amplifier and is fed through a special wide-dynamic-range amplifier to the quadrature video detectors.

The two quadrature video detector outputs are converted to 10-bit digital numbers by the A/D converters. The system contains an 8000-word input memory and about 900 integrated circuits. A disc memory is used as a fine-grained ground clutter map. This memory allows the removal of such returns by the use of adaptive thresholds in range, doppler and angle.

It is required that a number of scan returns be observed before a target is declared present and track is initiated. Thus, track initiation is used to help eliminate false clutter returns. Multiple PRFs are used to detect weak

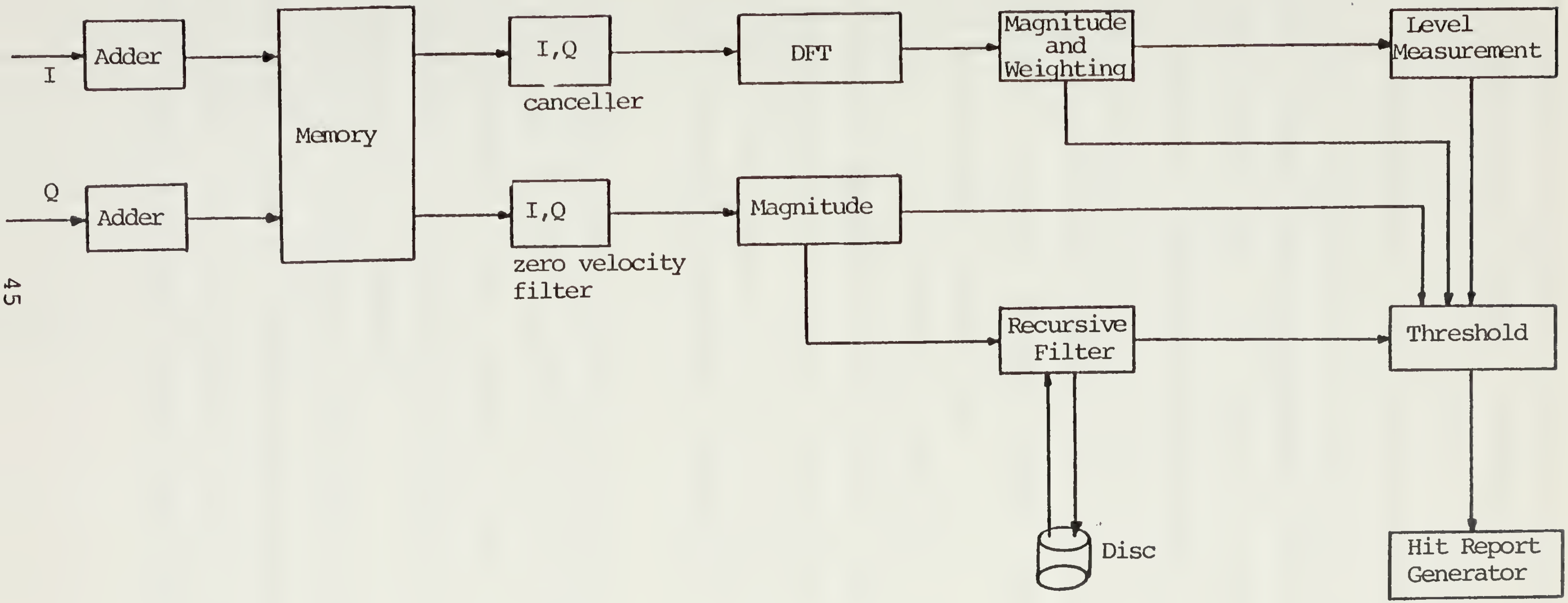


Figure 15. Moving Target Detector Processor

targets that ordinarily would be masked by rain clutter having the same ambiguous doppler frequency.

Because ground clutter is so spotty in nature, great difficulty is experienced in calculating appropriate threshold values for detection. To solve this problem a digital ground clutter map is implemented with one word for each range-azimuth cell. All these are stored on the magnetic disc memory. Sometimes magnetic-bubble and EBAM (Electron-Beam-Addressed Memory) are used, both of which have the advantage of being nonvolatile. Also, CCD memories are used, in place of disc systems, which are suitable for storing the clutter map.

MTD is said to have subweather visibility. In fact it can see a small commercial type aircraft whose cross-section is many dB below the radar cross-section of the weather return. This feature was not previously available. Now it is used to provide excellent tracking of aircraft in rain.

A second feature of the filter bank approach is the facility with which proper thresholds can be established taking into consideration the presence of rain. The detection threshold is established by summing the detected output in range cells of interest.

The MTD has the advantage of being able to detect and track zero doppler aircraft targets while ignoring bird and insect targets. At zero doppler, aircraft targets are broad-side and hence have much larger cross sections than

do birds or insects. Therefore, a higher threshold can be used in a zero doppler filter to eliminate bird echoes but still detect aircraft targets.

Tracking is automatically initiated and the tracks are continuous despite ground and weather clutter and birds. The radar now can be sited freely without consideration to ground clutter limitations.

Both MTI and pulse doppler processing have the disadvantage in regard to vulnerability to jamming. For both techniques many more than one pulse detection is required to see a signal imbedded in clutter.

Finally, no tuning or adjustment of any kind is needed for this combination. It is a reasonably economical solution to the radar automation problem.

C. FFT APPLICATION

Within the past several years, digital processing of radar signals has become quite popular. Recent advances in A/D converters, integrated technology, and the emergence of the Fast Fourier Algorithm have made practical the digital filtering of signals.

The application of the FFT algorithm starts from the first known technique of Convolution by the Discrete Fourier Transform. It is shown in Figure 16, where a block diagram explains the process.

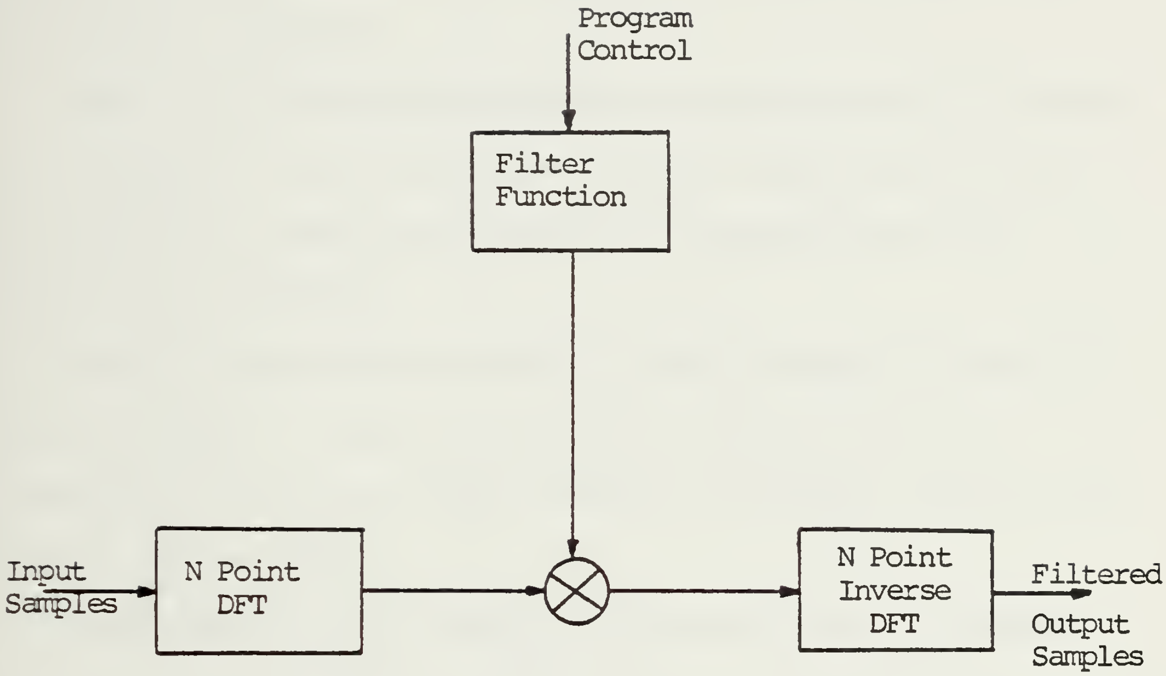


Figure 16. Convolution by discrete Fourier Transform

It is simple in concept and consists of taking an input sequence of samples from the received signals. Then a Fourier Transform is performed to yield samples in the frequency domain, multiplying these by a stored filter function and performing an inverse transform, yields the time domain response.

Thus, any filter function can be implemented by storing it in memory. So, knowing the clutter spectrum one can, in principle, filter the clutter by implementing the appropriate function.

The FFT algorithm is shown in the pipeline of Figure 17. This technique is used to digitally realize a bank of contiguous filters and thus produce a spectrum analyzer or doppler filter bank.

Advantages of the system include stability, reproducibility, and flexibility. The amount of time for this processing is a parameter of considerable interest because real time solutions are commonly required. To reduce time, high-speed A/D converters are being developed and employed in modern systems.

Following the discussion of the several techniques described in Chapters II, III and IV, a tabulation in matrix form comparing filter characteristics with clutter characteristics has been prepared and is shown in Table I. The classifications of techniques are qualitatively identified as Superior, Intermediate and Inferior, in terms of the amount of clutter suppression or improvement factor in

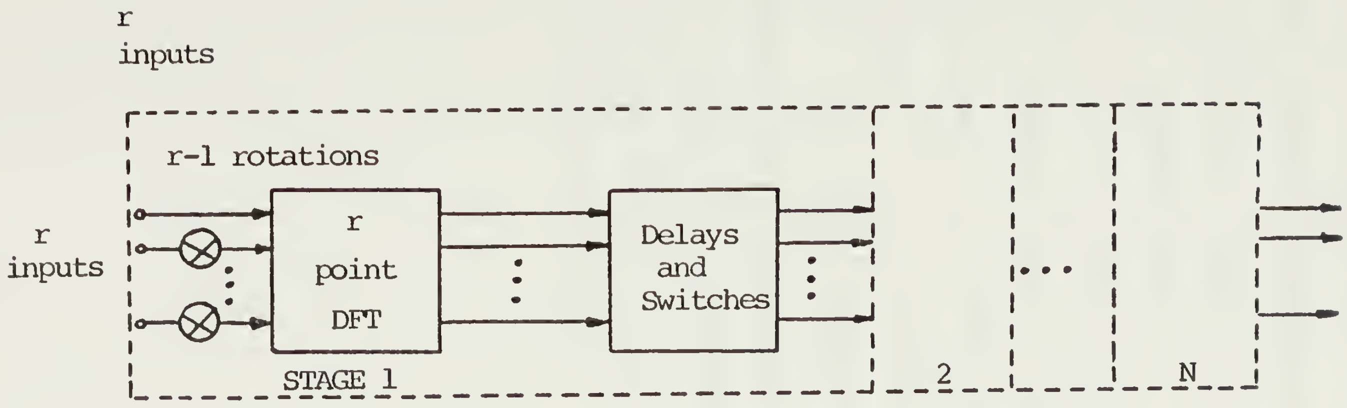


Figure 17. Pipeline FFT

dBs and also the range at which each process is reasonably sensitive where data information is available.

The purpose of this comparison is to help in making a decision about the system to be selected for a give use. It is noted that the decision will be based on the expected requirements from the system the area where it is to be installed, if a fixed or mobile site, and on other conditions which need consideration.

Each radar system is generally designed to optimize the realization of a specific set of measurement or observations. Once the design criterion has been achieved, usually it becomes difficult to utilize the radar for other purposes without extensive and costly modifications.

The observations and measurements shown in Table I were obtained from a number of independent observers. The data was taken chiefly at X-band on appropriately configured Navy and air-surveillance radars.

Technique / Clutter	Frequency Agility	Circular Polaris	STC-LOG RCVR	IAGC	MTI-Pulse Doppler	Dual Beam Antenna	LOG-CFAR	Video Detector	Pulse Dop Pr...	MTD	FFT
Ground	C 9-13 nm	B D	B 20 dB	B 15-20dB	B 30-40dB	B 15-20dB	B 20 dB	B	A 50-80 dB	A 52 dB 47.5 nm	A E
Sea	C 9-13 nm	B 10-20 dB	B 20 dB	B 15-20dB 5 nm	B 30-40dB	B 15-20dB R > 5nm	B 20 dB	B	A 50-80 dB	A 52 dB 47.5nm	A E
Rain	C	B 15 dB	B D	B D, 5nm	B D 13.8-24.6 dB	B	A 50 nm	B D	B	B	A E
Clouds	C	B 15 dB	C D	C	B D	B	A 50 nm	C	B	B	A E
Precipi-tation	C	B 15-20 dB	C	C	C D	C	B 15-20 dB 50 nm	C	B	B	A E
Birds Insects	C	C D	B D	B D	B D	C	B 15-20 dB	B	B	B	A E
Angels	C	C	B D,10 dB	C	B D	A D,15 dB	B	C	B	B	A E
Chaff	C 6 nm	B D	B D	B D	B D	B	B 15-20 dB	B	B	B	A E
Vegetation	C	C	C	C	C D	C	B	C	B	B	A E

Superior = A, Intermediate = B, Over 43 dB = E
Inferior = C, Short Range = D.

TABLE I. COMPARISON IN CLUTTER SUPPRESSION OF SELECTED TECHNIQUES

V. AN APPLICATION OF A DIGITAL FILTER

In the radars mentioned earlier, use of analog MTI systems give a clutter suppression of 13.8 to 24.6 dB for rain clutter according to the data of Table I. This suppression is not sufficient and the problem becomes greater when the radar system is operating under Mediterranean environmental conditions, where the fall and winter period is full of heavy rainfalls.

The experience of the Greek Navy for certain clutter due to sea and birds does not concur with the suppression values given in Table I. A level of 18 to 25 dB is achieved under perfect radiation conditions.

Since the radars under consideration are in use in the Greek Navy, a requirements exists for the improvement of their signal processors in order to get an improved clutter suppression.

This improvement can be achieved by the use of a higher than second-order doppler filter according to Ref. 12. This then is the motivation for the design of a fourth order recursive digital filter.

For this purpose a Butterworth characteristic bandpass filter is to be used with an effective bandwidth of 12616.63 rad/sec. Certain restrictions are put on the filter design as they are defined in the design part of this thesis.

The analysis and design are done according to Ref. 10. The concept of predestorted or prewarped analog transfer function is used. To get a one-to-one mapping in the z-domain approximately the same as that of the s-domain, the well-known bilinear z-transformation is used.

To detect moving targets, the system uses the doppler shift in frequency, which is given by the relation

$$f_d = \frac{2V_r}{\lambda} \tag{1.1}$$

The wavelength, λ, of the transmitted pulse might also be used as a parameter. Since the design of the above filter refers to specific systems, where λ is fixed by design parameter, this option is not open. These fixed values of λ for the two radars are calculated by using the relation

$$\lambda = \frac{c}{f_o} \tag{5.1}$$

where

$c = 3x10^8$ m/sec, velocity of light

$f_o =$ operating frequency of the system given in Hz

For f_o = 1300 MHz (AN/UPS-1) λ = .23 m;

and for f_o = 500 MHz (AN/SPS-40BC) λ = .6 m.

V_o stands for relative velocity of the target.

Usual types of moving targets, in the area of interest, are given in Table II with their cruise and maximum

Type	Cruise	Maximum
Helicopters Generally	80-130	180
C-47	150	250
C-130	200	300
ALBATROS	200	350
Commercial Air Lines	300-400	500
T-2E	250	600
A-7H	300	600
F-84F	350	600
F-5	400-550	700
F-102	500	800
F-104	400-550	1000
F-1CG	500	1000
F-4	500	1300

TABLE II. VELOCITIES OF SELECTED TARGETS

velocities in nautical miles per hour. This information was used in the beginning of the filter design in order to evaluate the necessary doppler frequencies assuming

(i) Radar stationary

(ii) Radar in motion with a velocity varying from 5 to 30 nautical miles per hour, to correct the effective bandwidth of the filter.

Table III contains a summary of the literature [Ref. 2, Chap. 6; Ref. 3, pp. 146, 551-554; Ref. 19, p. 17-9] for the relative velocities and standard deviations, σ_v, of clutter spectrum in meters per second. These kinds of clutter are common to the Mediterranean environment.

The conversion of the standard deviation of the clutter power spectrum, σ_c, to frequency is obtained by applying the relation

$$\sigma_c = \frac{2\sigma_v}{\lambda} \qquad (5.2)$$

where λ is the wavelength of the transmitted pulse in meters. This relation is an alternative to equation (1.1).

Computer programs 2, 3 and 4 were used to plot the amplitude square characteristic versus frequency, ω, of (1) analog, (2) prewarped, and (3) digital transfer functions. Figures 19, 20 and 21 show the characteristics of the designed filter.

Type of Clutter	Velocity Characteristics [m/sec]	Most Probable Cause
Vegetation:		
a. sparse woods	.017	calm condition
b. wooded hills	.04 - .32	winds with speeds from 10-15 knots
c. wooded hills	.01 - 3.0	winds with speeds up to 40 knots
Sea:		
a. sea echo	.7 - 1.0	droplet mechanism or facet motion
b. sea echo	.46 - 1.1	winds with velocity from 8-30 knots
c. sea echo	.89	windy condition
d. sea echo	.2 - 2.5	heavy windy conditions (to 30 knots)
Chaff:		
(i)	.37 - 1.1	
(ii)	1.2	Wind with velocity up to 25 knots
(iii)	4.0 - 6.0	Two different kinds of chaff. Experimental results by Medhurst, Jinsto and Gabie
Rain Clouds:		
a. thunderstorm (rain)	~9.0	heavy rains
b. normal	1.8 - 4.0	meteorological conditions
c. thunder snow	~1	meteorological conditions
Precipitation:		
(i)	~1	meteorological conditions

TABLE III. SUMMARY OF CLUTTER VELOCITY CHARACTERISTICS

Type of Clutter	Velocity Characteristics	Most Probable Cause
Angels:	Less than 85.72	meteorological effects, birds, insects, wind with velocity up to 20 knots, gravity waves, inversion of cold warm fronts, birds leaving roosting area, wind-carried echo at boundary of cloud
Birds, Insects:		
(i)	Less than 25.72	when carrying by wind during migratory periods (Spring-Fall)
(ii)	7.71 - 15.43	normal fly

TABLE III (Continued)

A. DIGITAL FILTER DESIGN

The second part of this thesis is the design of a digital filter which is going to reject clutter echoes in ranges indicated by Table III [Refs. 8,9,10,11,12]. Certain numbers of doppler filters combined with range gates constitute the video part of an MTI radar which is shown in Figure 18 in block diagram configuration.

A good performance criterion of MTI doppler filters is the MTI improvement factor. This factor defines the improvement of Signal to Clutter ratio which can be achieved by a filter. In the table of Ref. 12, the improvement factor I is given as a function of the order of the filter. Generally the higher the order of the filter the better the improvement factor. However, higher order transfer functions imply complexity, low reliability and higher cost.

The goal of the present doppler filter design is to derive a transfer function which will provide high clutter suppression and at the same time reduce the complexity caused by the higher order filter, to a reasonable level. For the above reasons, a fourth order digital filter will be designed. The improvement factor will be increased and the filter complexity will be kept to a minimum

An Infinite Impulse Response (IIR) band-pass filter of the fourth order will be implemented. As is known, the IIR filter does not need as many delays as the Finite Impulse Response (FIR) filter. Consequently, the number of weights is considerably lower.

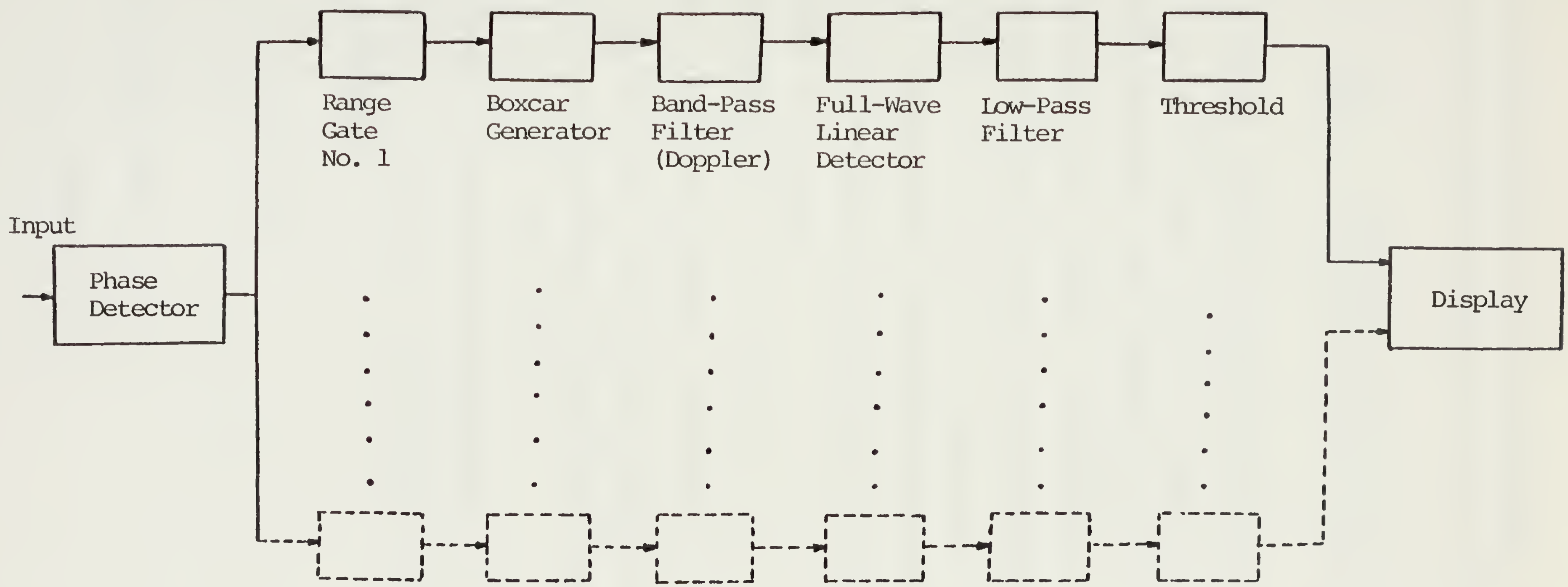


Figure 18. MTI using range gates and filters

The transfer function for a fourth-order band-pass digital filter is given by the following equation

$$H(z) = \frac{A(z)}{B(z)} = \frac{a_0 + a_1 z^{-1} + a_2 z^{-2} + a_3 z^{-3} + a_4 z^{-4}}{1 + b_1 z^{-1} + b_2 z^{-2} + b_3 z^{-3} + b_4 z^{-4}} \qquad (5.3)$$

This transfer function is completely defined by the values of a_0, a_1, z_2, a_3, a_4, b_1, b_2, b_3, b_4. There are two ways to define the above coefficients. The first is to go directly to the z-domain. According to the frequency response characteristics which will be defined later, one has to find the exact position of zeros and poles of H(z). Since zeros and poles are calculated the values of a's and b's occur after a simple multiplication of the factorized polynomials A(z) and B(z).

The second method, to be used here, is to select the desired H(s) transfer function. From this is defined the transfer function H(z). It is more difficult than the first method but has the advantage of permitting design of the filter in both the s and z domains.

The concept of predistortion applied to the transfer function H(s) is introduced. To transform H(s) to H(z) the z-transformation

$$z = e^{-sT_s} \qquad (5.4)$$

is normally used, where

T_s = sampling time.

Because of the cumbersome calculations resulting in the use of (5.4), the simpler bilinear z-transform

$$s = \frac{2}{T_s} \frac{z-1}{z+1} \tag{5.5}$$

is used. This is an approximation to (5.4)

The use of bilinear z-transform distorts the frequency response characteristics of $H(s)|_{s=j\omega}$. For this reason H(s) is predistorted in such a way that, after the bilinear transformation, H(z) will have the desired frequency response characteristics.

Starting in the s-domain a Butterworth fourth order band-pass filter is selected because of its flat response characteristics. The transfer function of such a filter is given by [Ref. 11]:

$$H(s) = \frac{bk\omega_o^2 s^2/Q^2}{s^4 + a\frac{\omega_o}{Q}s^3 + (2+\frac{b}{Q^2})\omega_o^2 s + a\frac{\omega_o^3}{Q}s + \omega_o^4} \tag{5.6}$$

where k = gain

$\omega_o = 2\pi f_o$ = center circular frequency

Q = quality factor which is given by $Q = \omega_o/BW$ and

BW = 3 dB bandwidth

$a = 1.41421$

$b = 1.0$

(constants given in Tables of Refs. 7 and 11)

If BW and ω_o are given, the frequencies at the 3 dB points, f_1 and f_2, can be defined:

$$\omega_2 - \omega_1 = 2\pi f_2 - 2\pi f_1 = BW \qquad (5.7)$$

$$\omega_1 \omega_2 = 2\pi f_1 \cdot 2\pi f_2 = \omega_o^2 \qquad (5.8)$$

The following restrictions are imposed on the filter design using charts.

(i) The maximum doppler frequency for clutter (angels, birds, insects) is 100 Hz. This corresponds to a velocity of 58.31 nm/h for λ = .6 m and 22.35 nm/h for λ = .23 m. The desired clutter suppression at 100 Hz is 30 dB.

(ii) The filter is to be used in air search radar and must detect targets with velocities of 881.7 nm/h for λ = .6 m and 340 nm/h for λ = .23 m. This sets the upper limit of frequency at 9500 rad/sec or 1572 Hz.

(iii) Slow moving targets such as helicopters must be detected with a suppression of no more than 3 dB. This sets the cut-off frequency at this end of the spectrum at 400 - 600 Hz.

With these restrictions the following values were selected.

ω_1 = 2829.44 4ad/sec, f_1 = 445.63 Hz

ω_2 = 15500 rad/sec, f_2 = 2463.7 Hz.

Then, applying basic relationships

$$\omega_o = 6622.47 \text{ rad/sec}$$

$$Q = .525$$

$$BW = 12616.63 \text{ rad/sec.}$$

With the constraints above and these values the function is prewarped and the bilinear z-transform applied. The final filter parameters were obtained after successive applications.

In Figure 20 the prewarped form of $|H(j\omega)|^2$ vs. ω is plotted. The prewarping was made on the center frequency ω_o with the exact Nyquist sampling time, $T_s = .2479445 \times 10^{-3}$ sec.

In Figure 21 the final filter power spectrum is given after the application of bilinear z-transform. Comparing this curve with the analog plot in Figure 19, it may be seen that a considerable distortion exists for frequencies $\omega > \omega_o$. This is reasonable as a result of the application of the transformation

$$\omega_a = \frac{2}{T_s} \tan \frac{\omega_D T_s}{2} \qquad (5.9)$$

where ω_a = analog circular frequency, rad/sec

ω_D = digital circular frequency, rad/sec.

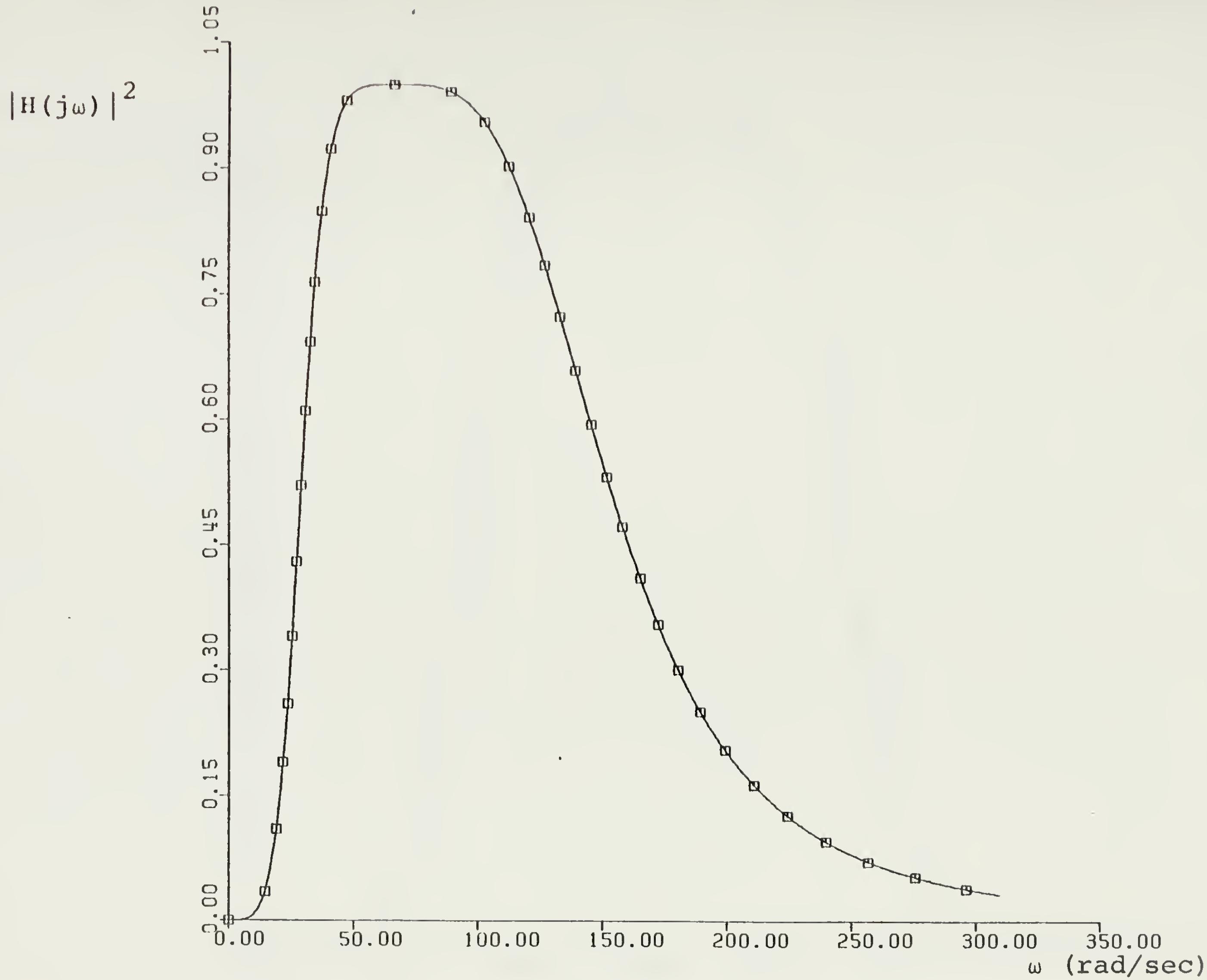


Figure 19. Amplitude characteristic of H(s)

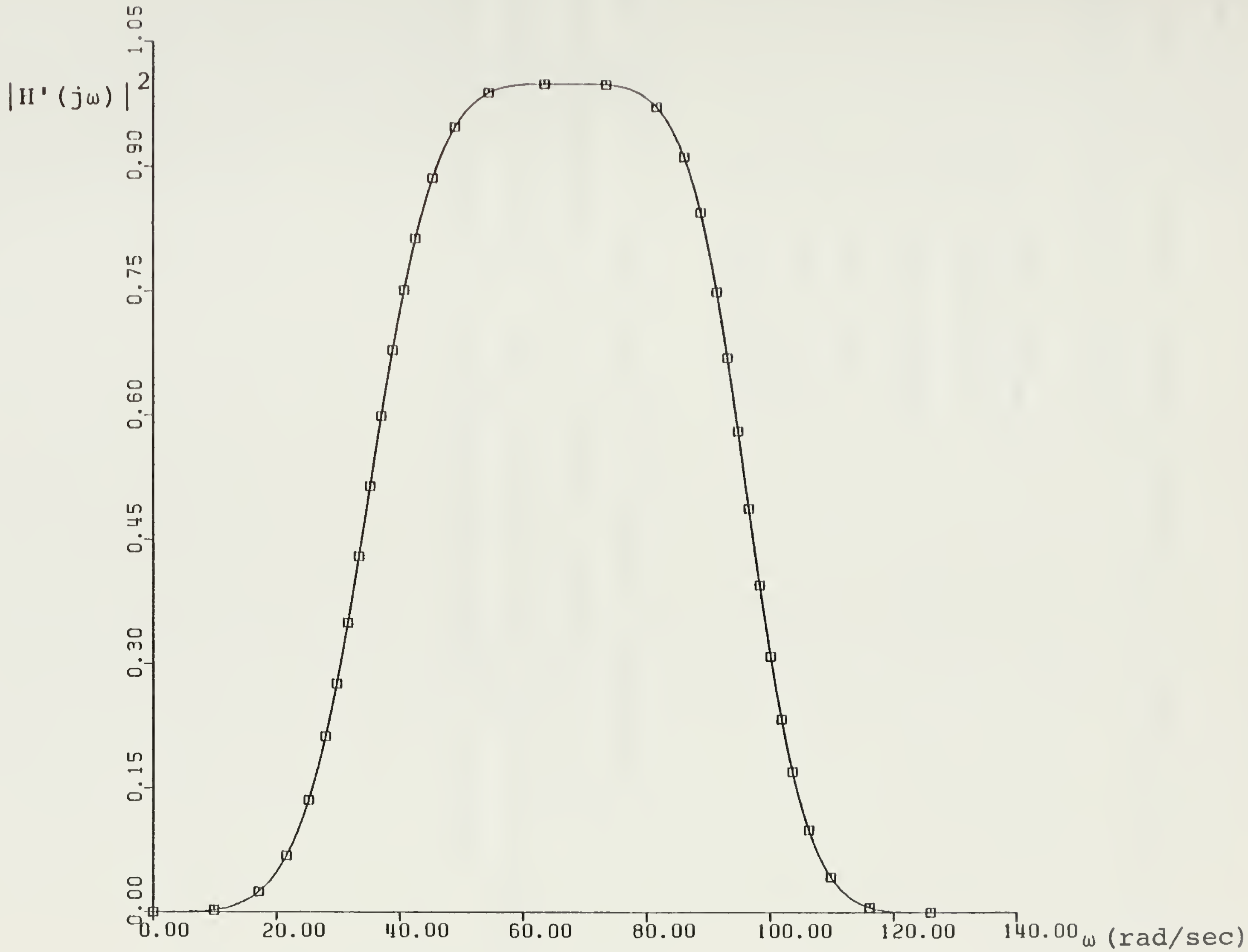


Figure 20. Amplitude characteristic of prewarped H(s)

For small values of ω_a, $\omega_a \cong \omega_D$, but for high values of ω_a, $\omega_a > \omega_D$. This is implied by the nonlinearity of equation (5.9).

The final filter design parameters are obtained from Figure 21 and are as follows:

$$\omega_2 = 9900 \text{ rad/sec}$$

$$f_2 = 1575.63 \text{ Hz}$$

$$\omega_1 = 3465 \text{ rad/sec}$$

$$f_1 = 551.47 \text{ Hz}$$

$$\Delta\omega = 6435 \text{ rad/sec}$$

$$\Delta f = 1024.16 \text{ Hz.}$$

More details of the analysis are given in Appendix A. Clutter suppression of 34.2 dB was calculated at the frequency of 100 Hz. This is better than that of the analog delay line which, in the existing MTI radar, is between 25 and 30 dB.

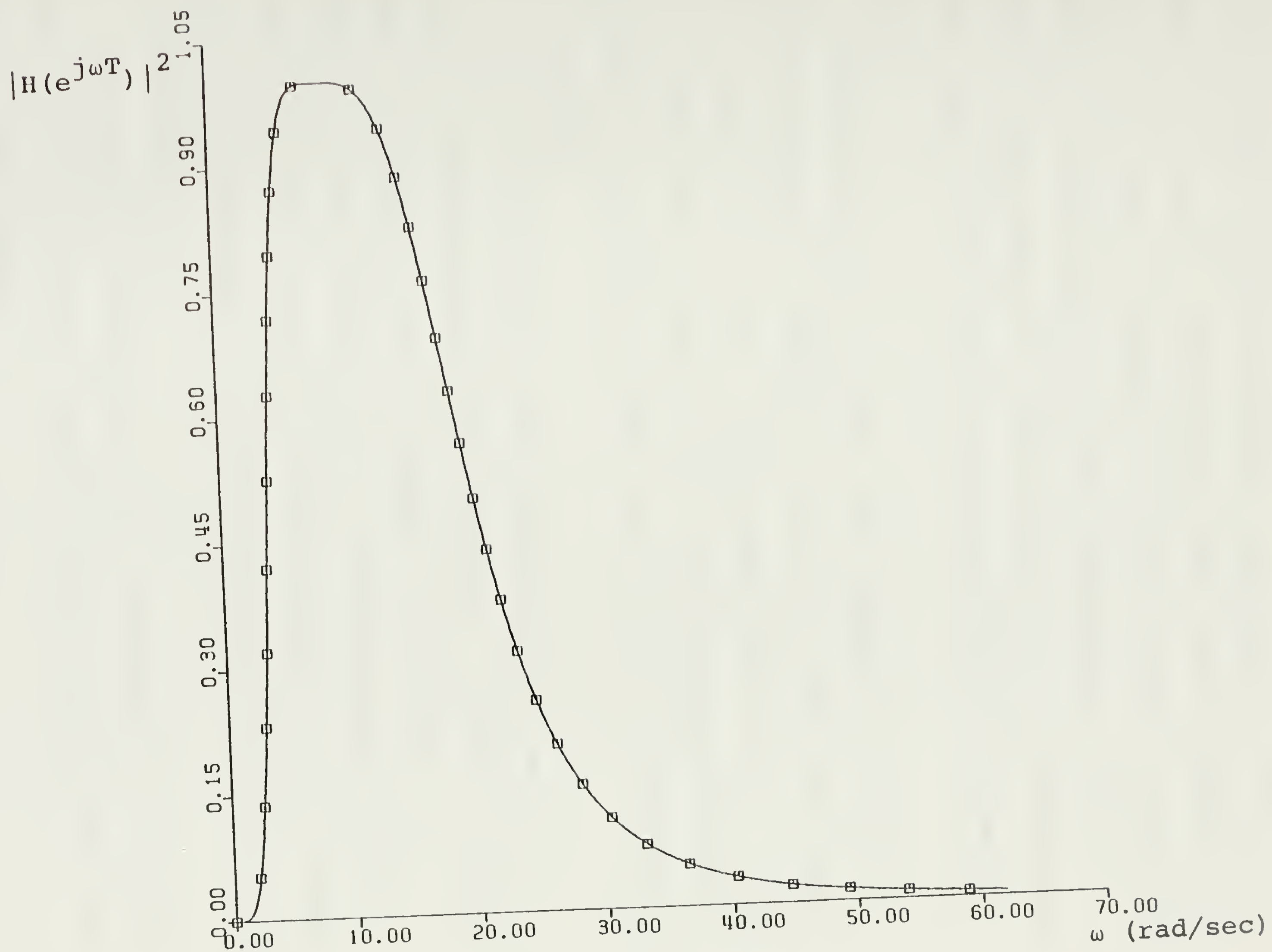


Figure 21. Amplitude characteristic of H(z)

VI. SUMMARY AND CONCLUSIONS

This thesis was motivated by a requirement to find an improved anticlutter technique for MTI radars presently used in the Greek Navy. Initially, a complete survey of existing methods of clutter suppression was undertaken. Table I summarizes the results of this survey showing qualitatively clutter improvement against various methods and types of clutter.

The Fast Fourier Transform (FFT) appears to be the optimum selection. However, it does not lend itself easily to introduction into the older radars still in use. There is no doubt FFT is used in modern radar systems and will be extended in the future, where cost and trained personnel are no obstacles.

For the application under consideration a method of digital signal processing was selected. It has the advantages of being practical, reliable and flexible. In addition, due to advances in LSI technology the digital filter may be contained in a small package of light weight with good performance characteristics and can be constructed at reasonable cost.

The design idea used here was introduced to the author for the first time in a digital filter course at the Naval Postgraduate School.

The concept of the prewarped transfer function was employed. This approach permits one to carry out his design in the familiar s-domain. The design constraints were taken from Table III. This table was assembled from a number of sources giving clutter characteristics for an unusual variety of clutter types. Since these -- birds, insects, angels and weather (squalls) -- are typical of the ones found in the Mediterranean environment against which the present radars do not give satisfactory suppression.

The transfer function H(s) was chosen from a table of filters of the Butterworth type. It is a fourth order filter. Knowing H(s), predistortion was applied in order to compensate for the distortion introduced by the application of the blinear z-transform. This is a powerful digital filter design tool since it permits the designer to control at each step the values of the filter parameters.

This thesis actually carried out the preliminary design of the selected filter. The results were then plotted using Continuous System Modeling Program. The filter parameters used in the program were those derived from the clutter characteristics surveyed as part of the thesis study. At the 100 Hz cut-off frequency this filter gives 4.2 to 9.2 dB improvement in clutter rejection. This is a significant improvement over the existing DLC MTI systems and is considered to have met the clutter requirements for ships operating in the Mediterranean environment.

The adaptive nature of the design proposed for the clutter problem suggests that it may be further utilized for other applications by interested students in the future. Additionally, digital filters may be actually constructed using the design program developed here and tested for correlation with the design parameters.

COMPUTER PROGRAM 1

Plot of clutter improvement factor vs. range

```
$$$CONTINUOUS SYSTEM MODELING PROGRAM III    V1M3   TRANSLATOR OUTPUT***

PARAMETER A=(.7,.75,.8,.85,.9,.95,1.,1.05,1.1)
       D1=1567.8
       D2=(A**2)+(.004*(R**2))
       DD=D1/D2
       R=TIME+3
TIMER FINTIM=120.,OUTDEL=1.,DELT=15.,PRDEL=15.
METHOD RKSFX
PRINT R,DD
OUTPUT R,DD(0.,2000.0)
             PAGE XYPLOT,WIDTH=10,HEIGHT=10
END
STOP
```

<u>COMPUTER PROGRAM 2</u>

Plot amplitude characteristic of analog H(s)

```
$$$CONTINUOUS SYSTEM MODELING PROGRAM  III    V1M3   TRANSLATOR OUTPUT$$$
        AA=-159179750.*F*F
        BB=0.
        CC=F*F*F*F-246894159.6*F*F+1923454500000000.
        DD=-17842.58*F*F*F+762525950000.*F
       F=TIME
        Y1=((AA*AA)+(BB*BB))**.5
        Y2=((CC*CC)+(DD*DD))**.5
        HJW1=Y1/Y2
        HJW=HJW1*HJW1
TIMER  FINTIM=45000.,OUTDEL=90.,DELT=310.,PRDEL=90.
METHOD RKSFX
PRINT  HJW,F
OUTPUT F,HJW
            PAGE XYPLOT,HEIGHT=7.,WIDTH=7.,
END
STOP
```

COMPUTER PROGRAM 3

Plot amplitude characteristic of prewarped H(s)

```
$$$CONTINUOUS SYSTEM MODELING PROGRAM  III   V1M3   TRANSLATOR OUTPUT$$$
        AA=-27222986.1*F*F
        BB=0.
        CC=F*F*F*F-42229661C.8*F*F+(56300055.E+08)
        DD=-23333.*F*F*F+(17507987.E+05)*F
       F=TIME
        Y1=((AA*AA)+(BB*BB))**.5
        Y2=((CC*CC)+(DD*DD))**.5
        HJW1=Y1/Y2
        HJW=HJW1*HJW1
TIMER  FINTIM=16850.,OUTDEL=90.,DELT=310.,PRDEL=90.
METHOD RKSFX
PRINT  HJW,F
OUTPUT F,HJW
           PAGE XYPLOT,HEIGHT=7.,WIDTH=7.,
 END
STOP
```

COMPUTER PROGRAM 4

Plot amplitude characteristic of H(z)

```
$$$CONTINUOUS SYSTEM MODELING PROGRAM  III    V1M3   TRANSLATOR OUTPUT$$$

CONSTANT K=.0002479445
         AA=COS(4.*K*F)-2.*COS(2.*K*F)+1.
         BB=SIN(4.*K*F)-2.*SIN(2.*K*F)
         PP=(35.95869*COS(4.*K*F))+(5.27084*COS(3.*K*F))
         LL=(2.38614*COS(2.*K*F))+(1.03484*COS(K*F))+6.19269
         CC=PP+LL
         QQ=(35.95869*SIN(4.*K*F))+(5.27084*SIN(3.*K*F))
         WW=(2.38614*SIN(2.*K*F))+(1.03484*SIN(K*F))
         DD=QQ+WW
        F=TIME
         Y1=((AA*AA)+(BB*BB))**.5
         Y2=((CC*CC)+(DD*DD))**.5
         HJW1=Y1/Y2
         HJW=99.9475275*(HJW1*HJW1)
TIMER FINTIM=18850.,OUTDEL=90.,DELT=310.,PRDEL=90.
METHOD RKSFX
PRINT HJW,F
OUTPUT F,HJW
             PAGE XYPLOT,HEIGHT=7.,WIDTH=7.,
END
STOP
```

APPENDIX A

Analysis of Digital Filter Design

This appendix covers the analysis in detail of the digital filter. The frequency domain transfer function is given in Ref. 11, pge 10-4.

$$H(s) = \frac{\omega_o^2 bks^2/Q^2}{s^4 + \frac{\omega_o}{Q} s^3 + (2 + \frac{b}{Q^2}) \omega_o^2 s^2 + \omega_o^4} \qquad (A.1)$$

where

$k = 1$, gain

$a = 1.41421$
(constants from Table 1, p. 10-4 of Ref. 11)
$b = 1.0$

$\omega_o = 6622.47$
$Q = .525$ (defined in the analysis)
$BW = 12616.63$

Applying these values to (A.1)

$$H(s) = \frac{1.59\text{x}10^8 \; s^2}{s^4 + 1.784\text{x}10^4 s^3 + 2.468\text{x}10^8 s^2 + 7.824\text{x}10^{11} s + 1.923\text{x}10^{15}} \qquad (A.2)$$

Applying $s \rightarrow j\omega$ (A.2)

$$H(j\omega) = \frac{-1.591\text{x}10^8 \; \omega^2 + j0}{(\omega^4 - 2.468\text{x}10^8 \omega^2 + 1.923\text{x}10^{15}) + j(-1.784\text{x}10^4 \omega^3 + 7.823\text{x}10^{11} \omega)}$$

This function was programmed in the computer and the results plotted in Figure 19. Associated program was Program Number 2.

According to the prewarping concept, the center circular frequency ω_o is prewarped. Since BW = 12616.63 rad/sec, then Δf = 2008 Hz, and

$$T_s = \frac{1}{2\Delta f} = .2479445\text{x}10^{-3} \text{ sec} \qquad (A.4)$$

The distortion factor ω' is given by the relation

$$\omega' = \frac{2}{T_s} \tan \frac{\Omega T_s}{2} \qquad (A.5)$$

But predistortion means

$$s \rightarrow \frac{s\omega_o}{\omega'} \qquad (A.6)$$

Applying (A.6) to (A.1) and substituting the known values

$$H'(s) = \frac{(\omega'/Q)^2 s^2}{s^4 + 1.41421(\frac{\omega'}{Q})s^3 + (2+\frac{1}{Q^2})(\omega')^2 s^2 + \frac{1.4142}{Q}(\omega')^3 s + (\omega')^4} \qquad (A.7)$$

Substituting (A.4) to (A.5) becomes

$$\omega' = 8662.18 \qquad (A.8)$$

Since Q = .525, (A.7) becomes

$$H'(s) = \frac{2.72\text{x}10^8\ s^2}{s^4 + 2.33\text{x}10^4 s^3 + 4.2\text{x}10^8 s^2 + 1.7\text{x}10^{12} s + 5.63\text{x}10^{15}} \qquad \text{(A.9)}$$

Applying $s \to j\omega$ (A.9) implies

$$H'(j\omega) = \frac{-2.72\text{x}10^8 \omega^2 + j0}{(\omega^4 - 4.2\text{x}10^8 \omega^2 + 5.63\text{x}10^{15}) + j(-2.33\text{x}10^4 \omega^3 + 1.7\text{x}10^{12} \omega)} \qquad \text{(A.10)}$$

This function was programmed to the computer and results in Figure 20. Associated program was Program Number 3.

Applying the bilinear z-transform, which is given by the relation

$$s \to \frac{2}{T_s} \ \frac{z-1}{z+1} \qquad \text{(A.11)}$$

to (A.9) becomes

$$H(z) = \frac{z^4 - 2z^2 + 1}{35.95869z^4 + 5.27084z^3 + 2.3814z^2 + 1.03484z + 6.19269} \qquad \text{(A.12)}$$

Finally, this function was programmed in the compuer and the results are plotted in Figure 21. Associated program was Program Number 4.

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